The proposed filter structure is shown in Fig. 2. To have an attenuation pole at 1.2 GHz, electromagnetic coupling between two resonators is controlled and capacitance patterns are added. The designed inter-resonators coupling constant is $k = 0.129$. Fig. 3 compares the simulated and measured results of the proposed filter. The 3-D field simulation is carried out with HFSS (Ansoft Co., Ltd.). The measured results show good agreement with the simulation. The measured passband insertion loss is 0.9 dB at the passband and the attenuation at 1.2 GHz is 60 dB and at 5 GHz is 70 dB. This filter is distinctive in the feature of its attenuation pole positions and this great performance is very favourable for wireless and Bluetooth applications.

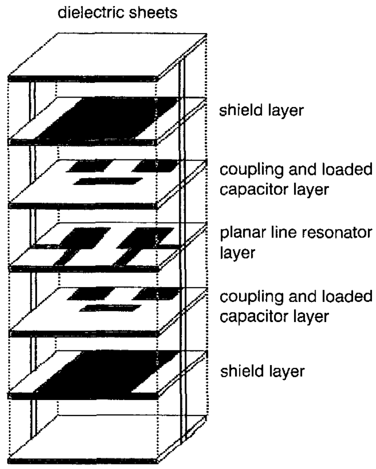

Fig. 2 *Filter structure*

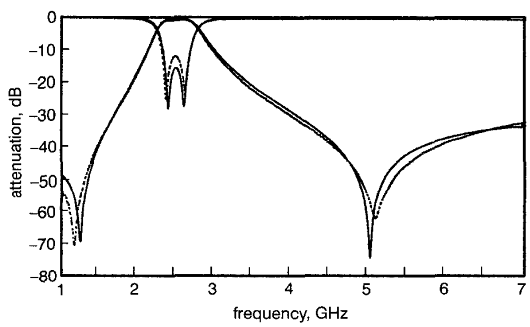

Fig. 3 *Response of proposed filter*

Conclusions: A novel laminated filter using tapped resonators is proposed which has attenuation poles at both sides of passband. Minimum 50 dB attenuations are achieved at the attenuation pole positions. This performance is applicable to wireless and Bluetooth applications.

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B.H. Lee, S.S. Park and J.H. Yoon (*CAE Team, R&D Center, Samsung Electro-Mechnics Co., Ltd, 314, Maetan-3Dong, Paldal-Gu, Suwon, Kyunggi-Do, 442-743, Korea*)

D.S. Park (*MLCC Lab, Electronic Device Business, Samsung Electro-Mechnics Co., Ltd, 314, Maetan-3Dong, Paldal-Gu, Suwon, Kyunggi-Do, 442-743, Korea*)

E-mail: as89@samsung.co.kr

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Phase noise in distributed oscillators

C.J. White and A. Hajimiri

The phase noise of a distributed oscillator is evaluated very simply by identifying an effective capacitance equal to the total capacitance distributed along the transmission lines. The contributions of the various passive and active noise sources to the total phase noise are calculated revealing several guidelines for improved distributed oscillator designs.

Introduction: Distributed oscillators have various applications including multi-gigahertz integrated voltage-controlled oscillators [1, 2] and broadly tunable microwave oscillators [3]. Calculating the phase noise of a distributed oscillator appears to be a much more complicated problem than the lumped case; however, we demonstrate that once the basic equation for phase noise in distributed oscillators is established, the techniques for calculating phase noise in lumped circuits [4–6] are easily adapted to the distributed case.

One example of a distributed oscillator is the forward gain mode oscillator shown in Fig. 1. The oscillator consists of two transmission lines connecting the gate and drain of several evenly spaced transistors. The output at the right of the drain line is AC coupled to the input at the left of the gate line. Matched terminations on the opposite ends of each transmission line absorb the reverse waves. The operation of the oscillator is given in more detail in [1].

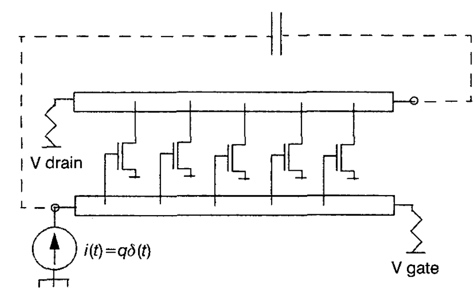

Fig. 1 *Forward mode distributed oscillator (biasing not shown)*

Einstein relation: The conjecture behind the results presented in this Letter is based on the phase noise theory developed in [6]. The phase noise of a lumped LC oscillator can be calculated using the Einstein relation for the phase diffusion constant [6]. For a sinusoidal oscillator, the phase diffusion due to passive, white noise sources is

$$D = \frac{1}{2} \cdot \frac{kT}{CV_o^2} \cdot \frac{\omega_o}{Q} \tag{1}$$

In (1) C is the tank capacitance, V_o is the oscillation amplitude, k is Boltzman's constant, T is temperature, ω_o is the centre frequency and Q is the loaded quality factor of the resonator. The factor of 1/2 accounts for time-varying effects. The corresponding phase noise is [6]

$$L\{\Delta\omega\} = 10 \log\left(\frac{2D}{(\Delta\omega)^2}\right) \tag{2}$$

For the distributed oscillator, we will show later that the lumped capacitance C is replaced by the total capacitance distributed along the transmission lines. The Q of the distributed oscillator is calculated using the general definition (ω_o times the energy stored in the oscillation divided by the average power dissipated). For the oscillator in Fig. 1 Q is essentially π, because the sum of the power dissipated in the gate and drain lines and the power lost to the matched terminations does not depend on the attenuation coefficient of the lines.

Langevin equation: The detailed calculation of the phase noise of a distributed oscillator is established by deriving the Langevin equation for the phase diffusion [4]. Consider a current impulse $q\delta(t)$ injected at the input of the distributed oscillator, as shown in Fig. 1. The voltage at the input is the sum of the oscillation (with amplitude V_o)

and the impulse (represented as a Fourier series):

$$v_{\text{in}}(t) = V_o \cos\Big(2\pi\Big(\frac{v}{2nl}\Big)t + \theta\Big) + \sum_{n=0}^{\infty} q \frac{Z_o v}{2nl} \cos 2\pi\Big(\frac{(2n+1)v}{2nl}\Big)t \quad (3)$$

In this expression, Z_o is the characteristic impedance, v is the phase velocity, and nl is the total length of each transmission line. The voltage along the gate line is similar to the waveform in Fig. 2, where the dashed line shows the sum of the oscillation and the fundamental component of the injected current. The phase shift is

$$\Delta\theta = \frac{(q(Z_o v/2nl))}{V} \sin\theta \quad (4)$$

assuming that the amplitude of the fundamental component ($qZ_o v/2nl$) is much smaller than the oscillation. The parasitic capacitances of the transistors filter the higher harmonics of the impulse, so that the total amplitude of the noise signal at the gate of the transistors is small. To first order, the injected noise does not change the operating point of the oscillator, except for the phase perturbation calculated above.

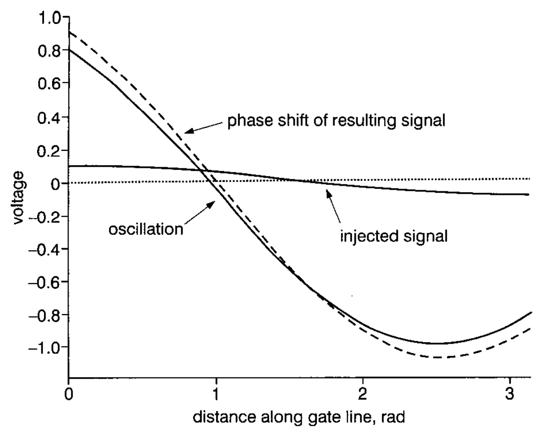

Fig. 2 *Voltage along gate line, showing oscillation and fundamental component of injected impulse*

Comparing the Langevin equation for the phase of a distributed oscillator, (4), to the standard equation for a sinusoidal oscillator (c.f. [4]), the equations are the same except that the capacitance of the tank circuit is replaced by the total capacitance $C_T = 2nl/vZ_o$ distributed along the transmission lines. The phase noise of a distributed oscillator in response to a white noise current source with power spectral density (PSD) $i_n^2/\Delta f$ injected at the input is therefore

$$L\{\Delta\omega\} = 10 \log\left(\frac{1/2}{(2nl/Z_o v)^2 V^2} \frac{i_n^2/\Delta f}{2\Delta\omega^2}\right) \quad (5)$$

in the $1/f^2$ region, where the factor of $1/2$ is the average of the impulse sensitivity function (ISF) for a sinusoidal oscillation [5].

Noise sources: For the distributed oscillator shown in Fig. 1, the thermal noise from the drain line and left termination is always $4kT/Z_o$. This can be calculated from the usual Nyquist equation for the thermal noise of a resistor, because the impedance of the matched transmission line is Z_o. As a result, the phase noise of the oscillator must be greater than

$$L\{\Delta\omega\} > 10 \log\left(\frac{kT}{V^2} Z_0 \frac{f_o^2}{\Delta\omega^2}\right) \quad (6)$$

For other designs, it may be useful to calculated the noise generated by the transmission line only [7, 8].

Thermal noise generated by the gate line adds to the total phase noise of the distributed oscillator in Fig. 1. The details of the calculation are omitted.

Noise from the transistors is cyclostationary, and needs to be adjusted by ($\Gamma^2_{eff,rms}$), the *rms* value of the product of the noise modulation function (*NMF*) and the impulse sensitivity function (*ISF*), as discussed in [5, 6]. The noise power from each transistor that reaches the input is reduced by the power loss along the drain line, depending on the position of the transistor.

The overall phase noise for the distributed oscillator in Fig. 1 is therefore

$$\{\Delta\omega\} = 10 \log\left(\frac{Z_o^2 \sum \Gamma^2_{\text{eff,rms}} \cdot i_n^2/\Delta f}{V^2} \frac{f_o^2}{2\Delta\omega^2}\right) \quad (7)$$

where the sum is over all the passive and active noise sources, and $\Gamma^2_{eff,rms}$ is determined by the cyclostationary properties of the noise sources.

Conclusion: We have presented a general method for calculating the phase noise of a distributed oscillator. The total capacitance distributed along the transmission lines enters into the Langevin equation for the phase diffusion in place of the lumped capacitance of the resonator. To design an improved distributed oscillator, the power loss to the terminations must be eliminated to increase the overall Q. In the current design (Fig. 1), the phase noise is largely independent of the loss in the transmission lines.

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C.J. White and A. Hajimiri (*Department of Electrical Engineering, California Institute of Technology, Pasadena, CA 91125, USA*)

E-mail: cjwhite@caltech.edu

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Improvement of iterative decoding algorithm for linear block codes

G. Drolet

Two parameters are introduced into a popular algorithm for the iterative decoding of information encoded with a binary linear code and transmitted over the additive white Gaussian noise channel to improve the coding gain. Theoretical justification supported by simulation results are presented.

Introduction: Let $\mathbf{F}_2 \cong \text{GF}(2)$ denote the binary field and $\mathcal{C} \subset \mathbf{F}_2^n$ **a**(n, k, $d_{\min}$) binary linear block code with (n, $n-k$, $d^{\perp}_{\min}$) dual code $\mathcal{C}^{\perp} \subset \mathbf{F}_2^n$. One codeword $\boldsymbol{c} = (c_1, c_2, \ldots, c_n) \in \mathcal{C}$ is chosen at random among the 2^k equiprobable codewords of $\mathcal{C}$. The bits of $\boldsymbol{c}$ are transmitted over the (memoryless) AWGN channel using (normalised energy) binary antipodal modulation:

$$\mathbf{F}_2 \rightarrow \{\pm 1\} \subset \mathbf{R}$$
$$b \rightarrow (-1)^b$$